ORTHOGONAL SELF-ADAPTIVE CONTROL FOR MINIMISATION OF PHASE ERROR

By

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CERTIFICATE

Certified that the dissertation entitled "ORTHOGONAL SELF-ADAPTIVE CONTROL FOR MINIMIZATION OF PHASE ERROR" which is being submitted by Sri S.K. Gupta in partial fulfilment for the award of Master of Engineering in "Applied Electronics and Servomechanism" of University of Roorkee is a record of student's own work carried out by him under my supervision and guidance. The matter embodied in this dissertation has not been submitted for the award of any other Degree or Diploma.

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- SYNOPSIS -

Automatic phase control (APC) systems, which are basically non-linear feed back control systems, have been analysed for different inputs using different techniques[8,9,10,11]. The APC systems normally do not respond to amplitude modulated inputs which impose limitations on the use of these systems in many cases.

In this thesis, the problem of neutralizing various errors in APC systems is discussed. The use of self adaptive control system is suggested and discussed in detail to achieve this end. The modification is made in the open loop system thereby preserving the basic characteristics of APC system. It has been shown that the modified system can be used both for A.M. and F.M. inputs without introducing various errors.

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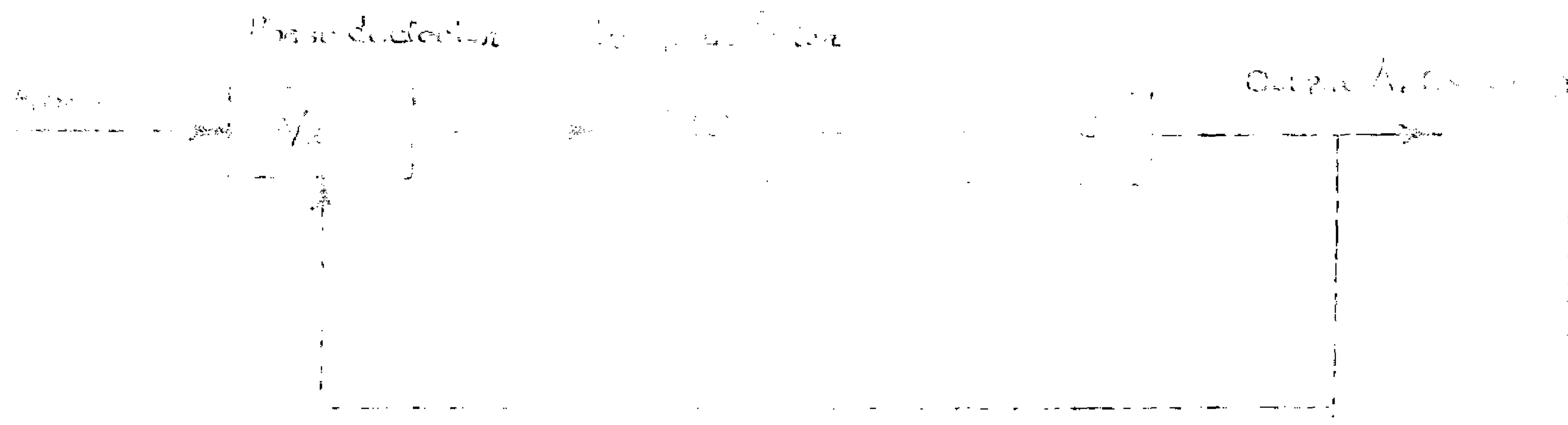
CHAPTER - I

1.1 INTRODUCTION :

Phase lock systems (also known as automatic phase control systems) are used as tracking filters, synchronising devices and as narrow band F.M. discriminatòrs. As a tracking filter, the A.P.C. system helps in increasing S/N ratio of a receiver; using different parameters , it can also act as accurate, sensitive, synchronising device for increasing the frequency stability of a high powered oscillator; as narrow band F.M. discriminater it finds important application in frequency measurement and synthesis.

The basic A.P.C. system is shown schematically in Fig. 1.1. It essentially consist of a phase detector and a voltage controlled oscillator. In addition a low pass filter is also used to improve the performance of the system. Low pass filter may be a simple R.C. filter, but usually a more complex lag network is used. The lag network helps to increase the capture range of the system. If it is desired to filter the noise and interference from the input or if a highly stable local oscillator operation is required a narrow band filter is used on the other hand if the input signal has to be tracked a wide band filter is to be employed.



[illegible] 1 [illegible] SYSTEM

A.P.C. system is a feed back control system and the error actuating the oscillator is phase error between input reference signal and the oscillator output. For constant frequency inputs oscillator output follows the input with a phase error , but as A.P.C. system is a type - I servomechanism for F.M. signals, the oscillator output follows the input with a frequency error. These lock systems do not respond to amplitude variations in the input signal and thus the amplitude information of the input signal is lost. In this thesis, it is described that using a self adaptive system it becomes possible to follow the input in phase, frequency and in amplitude.

1.2 BASIC OPERATION :

When the system is locked its operation is such that the oscillator phase adjusts itself in such a manner that the phase detector output is of the correct value for keeping the oscillator frequency equal to the input signal frequency. As the input signal frequency varies , the phase difference varies accordingly. If input frequency is exactly equal to the natural frequency of the oscillator, the actuating error then ϕ will be zero and hence the phase difference ϕ between input signal and oscillator output will be 90°. Depending upon the magnitude of the mistuning and the input amplitude the oscillator phase differs from 90°. The limit is reached when it is 0° or 180°. The phase detector voltage can not be increased any further by changing oscillator phase and hence beyond this value of mistunning the synchronization fails. This range of frequency over which the system can remain in synchronism is called the lock range, which is equal to $2 K_1 K_2$ (1) where K_1 is the phase detector gain in volts/radian and K_2 is the reactance tube sensitivity in radians/sec/volt.

The addition of the filter reduces the capture range, that is the frequency range within which the oscillator synchronises with the input. Once the oscillator is locked in synchronism it will remain locked till the extremities of the lock range.

2.3 SYSTEM RESPONSE TO CONSTANT FREQUENCY INPUT:

Basic differential equation of the system is derived in Appendix - I, equation - 1. It shows that the system performance depends upon F(s) i.e. filter transfer function.

Steady state solution of the equation will be

$$\sin \phi = \frac{\Omega}{K \; F(o)}$$

where Ω = is detuning to be locked in i.e. the difference of input and oscillator natural frequency

K = is the system gain

ϕ = is the phase error between input and oscillator frequency when the oscillator is locked to input frequency.

As once the system is locked phase detector output is d.c. voltage for constant frequency inputs. So the steady state error will be :

$$\phi_{ss} = \text{Arc sin} \frac{\Omega}{K}$$

(a) Filter less case :-

If no filter is used the system behaves like a simple R.C. low pass filter having transfer function

$K/s + K$

where K is system gain cut off frequency of the filter, will be $\omega = K$.

If no filter is used the capture range of the system is equal to the lock range of the system, so for any input having frequency within $\pm K$ radian/sec of the natural frequency, the oscillator will synchronise to it. But as the detector output is a sine function of phase difference between input and oscillator output so a little perturbation at extreme points will derive the oscillator to unlock.

Transient response of the system to a step input is given by equation - 9, Appendix II which is an increasing exponential function like an over damped system.

(b) With Filter :-

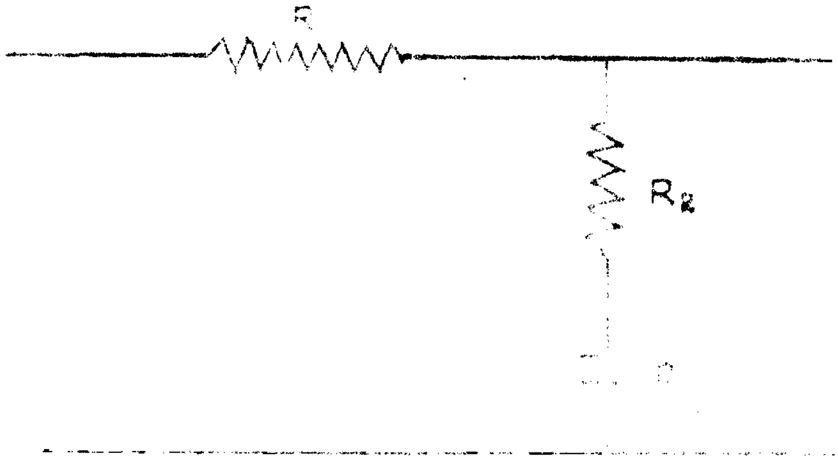
Transfer function of a low pass filter is ;

$$F(s) = \frac{1 + a \tau s}{1 + \tau s}$$

For a = 0, it is a simple R.C. filter and for a = 1, filter transfer function is unity and it is a filterless case.

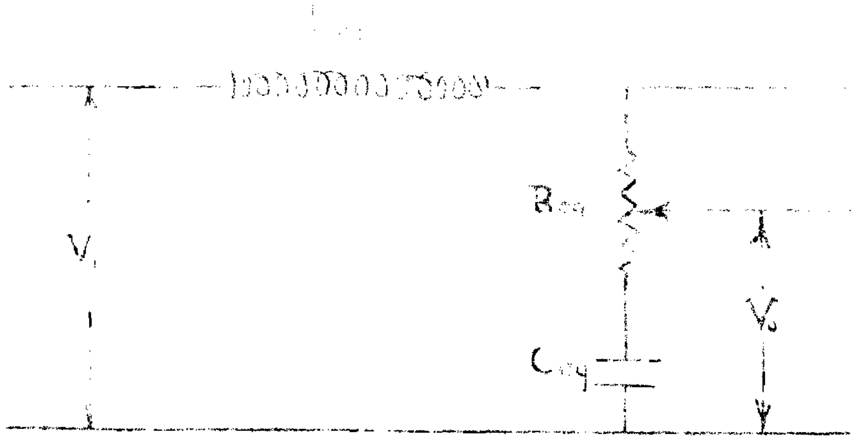
With a filter having transfer function $F(s) = 1 + a\tau s/1 + \tau s$ as shown in Fig. 1.2, the system equation (equation 15, Appendix - II) is a non linear second order differential equation but for small change in phase difference the system equation can be linearised.

The locked system with compensated R.C. filter can be represented by a passive network[2] shown in Figure - 1.3 .



$$F(s) = \frac{1 + a\tau s}{1 + \tau s}$$

Fig 1.2 Lag Network



$$L_{eq} \cdot C_{eq} = \frac{1}{\omega_n^2}$$

And $\frac{R_{eq}}{L_{eq}} = 2\xi\omega_n$

Fig 1.3. Equivalent Passive Network for APC System

But due to non linearity the output and the network parameters will be function of ϕ, the steady state error,

$$\text{Leq.}\ C_{eq}^{(2)} = \frac{1}{\omega_n^2} = \frac{\tau}{K \cos\phi}$$

$$\frac{R_{eq}}{L_{eq}} = 2\ \zeta\omega_n = \frac{(1 + a K \cos\phi)}{\tau}$$

and the output will be $V_o \cos\phi$ instead of V_o.

Frequency response of the system for small change in ϕ is given by equation - 10, Appendix II and is shown in Figure. 1·4

Transient response :-

Transient response for linearised system is shown in Figure - 1.6 for value of $\zeta = 1/2$ and $\omega_n/K = 0$ and $\omega_n/K = 1$.

Capture Range :-

Capture range of the system is[2] (vide Appendix - II)

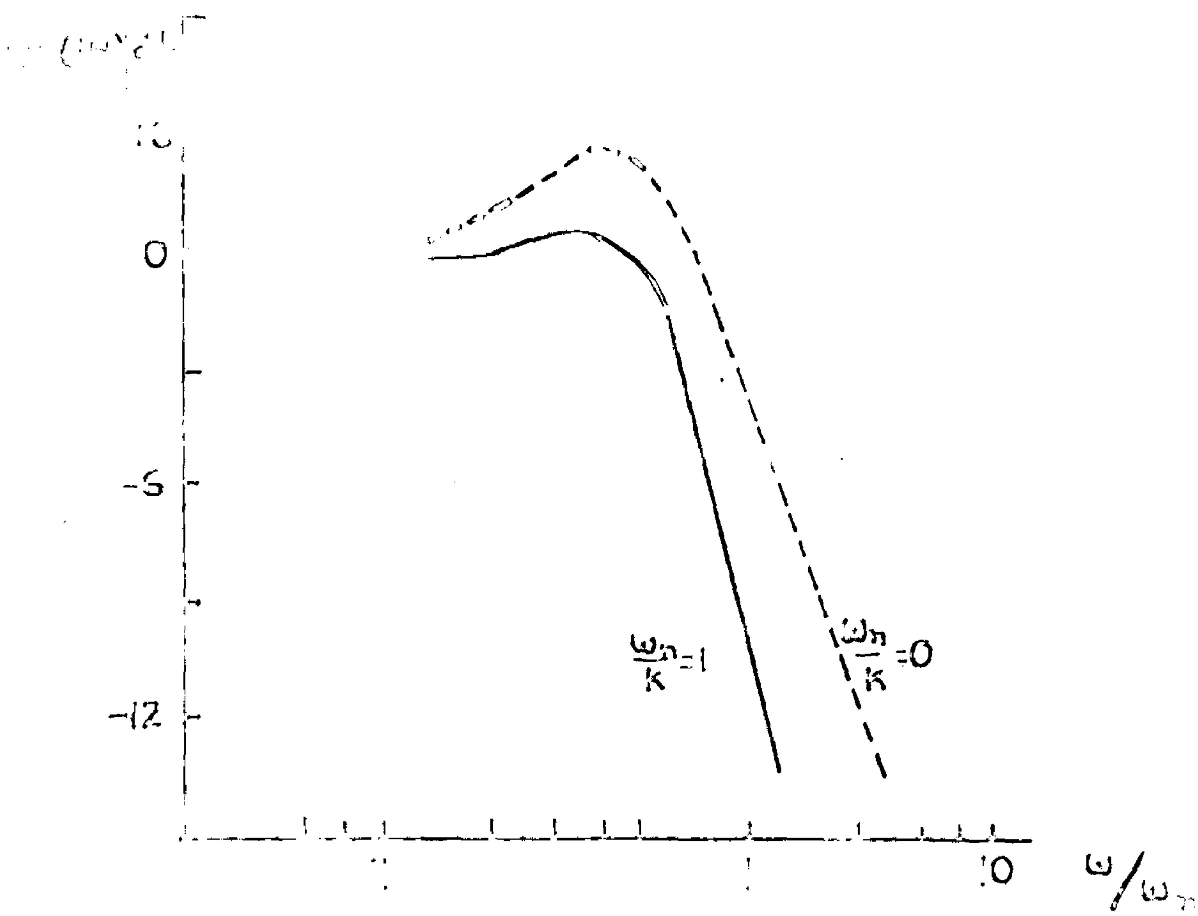
$$\sqrt{2\zeta\omega_n K}$$

where

ζ is damping coefficient

ω_n is natural frequency of the system

K is system gain.



FREQUENCY RESPONSE FOR $\zeta = 0{\cdot}5$

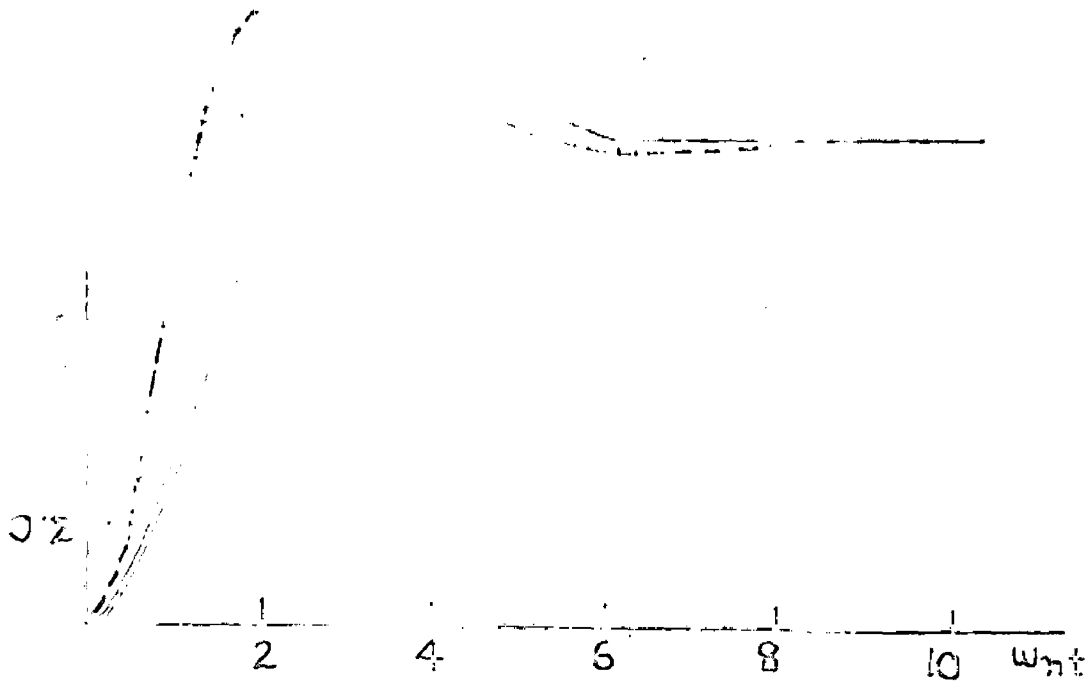


FIG 6: TRANSIENT RESPONSE FOR $\zeta = 0{\cdot}5$

For ξ equal to 1/2, the capture range is $\sqrt{\omega_n K}$ and the capture range to lock range ratio (Appendix - II) is $\sqrt{\omega_n / K}$ which is ω_n / K for simple R.C. filter. It shows that the reduction in capture range using a lag network is square root of the reduction in capture range using a R.C. filter for the same band width.

Steady state error will be :

$$\phi_{ss} = \sin^{-1} \frac{\Omega}{K}$$

where Ω = initial detuning

K = system gain.

If the initial detuning is more than capture range the system will not synchronise to the input but will exhibit hysteresis effect[3] .

A.F.C. SYSTEM RESPONSE TO F.M. SIGNAL :

If the system is assumed to be linear then it is a type - I system[4] and the system will follow the ramp modulated input with a constant frequency error given by :

$$\frac{\delta}{K F(j\omega)}$$

where δ = modulation index

K = system gain

$F(j\omega)$ = filter gain for error frequency.

But as the system is not linear, for $\delta/\omega_n^2 < 1/2$ the system will lock to input frequency and will remain locked, if once it is locked till $\delta/\omega_n^2 = 1$ (5). The phase error will change from 180° to 0°.

For sinusoidal modulating wave i.e. input frequency is

$$\omega_1 + \delta \sin \omega_m t$$

Now assuming $\omega_1 = \omega_o$ i.e. input centre frequency is equal to natural frequency of oscillator, error ϕ will function of time and as shown by equation 22, Appendix III.

$$\phi(t) = \sin^{-1} \left(\frac{\sin \omega_m t}{K F(j\omega)} \right)$$

where δ = is modulation index

ω_m = modulating frequency

K = system gain.

The oscillator output is not truly the input but it is following the input with error which is a function of modulating frequency and modulation index.

c. RESPONSE TO A.M. INPUT :

A.P.C. system output does not follow any amplitude information present in input. As the phase detector out put is a function of input amplitude . So for amplitude modulated signal the phase error ϕ will be function of input amplitude that is oscillator output will be phase modulated.

Maximum allowable modulation for which the oscillator will synchronise with input depends upon initial detuning. For system, to remain in synchronism the modulation should not be more than

$$m < 1 - \frac{\Omega}{K} \quad \text{(vide Appendix - III).}$$

Approximating for zero initial detuning, the system without filter will lock for maximum 100% modulated input.

LIMITATIONS :

a) Phase Error :-

A.F.C. system is a closed loop feed back system so to control the oscillator frequency. It is essential to generate some error function. Here the error is function of phase difference between input and oscillator output. Hence the system is not following the input in phase and all phase information in input are lost.

b) System Response to F.M. Input :-

As it is a type - I system so it follows the F.M. signal with a frequency error. For F.M. input signals the informations contained in the output are not the same as in input. Use of A.F.C. system as a tracking filter is restricted for the targets changing its velocity as it is not giving correct information.

Using one more integrator the system can be modified for the use of F.M. inputs. As now the system becomes type - II and the velocity error becomes zero.

c) Response to A.M. Signals :-

Out put of the A.F.C. system gives no amplitude information, so all the amplitude informations contained in input are lost. Simple A.F.C. system can not be used where the amplitude informations in the input are to be utilized.

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APPENDIX - I

BASIC DIFFERENTIAL EQUATION OF THE AFC SYSTEM.

Figure 1-(A) represents the loop configuration. Input reference signal frequency is Ω_1 may either constant or time varying.

VCO output frequency is oscillator centre frequency ω_c plus the time varying term proportional to actuating signal e_a. When the input reference signal frequency is equal to oscillator centre frequency ω_c, VCO output is 90° out of phase to input signal.

Phase detector out put

$$e_d = K_o \sin \phi \quad \text{- (1)}$$

where $\phi = \phi_2 - \phi_1$

actuating signal

$$e_a = K_a K_1 F(s) \sin \phi \quad \text{.......... (2)}$$

ϕ = phase difference between output and input signal

VCO output frequency will be ;

$$\omega_o = \frac{\omega_{oc} + K_o K_1 K_2 F(s)}{s} \sin \phi \quad \text{.....(3)}$$

Hence phase detector output will be equal to ,

$$e_d = K_o \sin \phi$$

$$= K_o \sin \left\{ \phi_s(t) - W_c t - \frac{KF(s)}{S} \sin \phi \right\}$$

where $k = k_o k_1 k_2$

$$\text{or} \quad \phi = \phi_s(t) - W_c t - \frac{KF(s)}{S} \sin \phi \qquad \text{....(4)}$$

$$\text{or} \quad \dot{\phi} + KF(s) \sin \phi = \dot{\phi}_s - W_c$$

$$\text{or} \quad \dot{\phi} + K\, F(s) \sin \phi = \Omega \qquad \text{......(5)}$$

where $\Omega = \dot{\phi}_s - W_c$

at steady state $\dot{\phi} = 0$. Hence the solution of equation (5) for steady state is ;

$$\sin \phi = \frac{\Omega}{K\, F(s)} \qquad \text{....(5.a)}$$

or $\sin \phi = \Omega / K$ as when the system is locked $F(s) = 1$.

$$\text{So} \quad \phi_{ss} = \sin^{-1} \frac{\Omega}{K} \qquad \text{...(6.b)}$$

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APPENDIX - II.

(a) Filter Less Case :-

When the filter transfer function is unity equation . 5 becomes

$$\dot{\phi} + K \sin \phi = \Omega \qquad ......(7)$$

It is a simple first order non linear equation

System transfer function is :

$$G(s) = \frac{K}{s + K} \qquad .....(8)$$

assuming ϕ small

so that $\sin \phi \approx \phi$

System behaves like a simple R.C. low pass filter; cut off frequency is $\omega = K$

Transient response to step input will be :

$$\phi_2(t) = V(t)(1 - e^{-Kt}) \qquad ......(9)$$

(b) With Filter :

When the filter transfer function

$$F(s) = \frac{1 + a\tau s}{1 + \tau s} \qquad ......(10)$$

Substituting in equation - 5,

$$\dot{\phi} + K\left(\frac{1 + a\tau s}{1 + \tau s}\right) \sin \phi = \Omega$$

or :

$$\ddot{\phi} + \left(\frac{1}{T} + aK\cos\phi\right)\dot{\phi} + \frac{K}{T}\sin\phi = \frac{\Omega}{T} \quad \ldots(11)$$

Equation (11) is a second order non linear differential equation. Its behaviour can be found out using phase plane technique[8].

For small phase error the equation (11) can be linearised to ;

$$\ddot{\phi} + \left(\frac{1}{T} + aK\right)\dot{\phi} + \frac{K}{T}\phi = \_\_\_\_ \quad \ldots\ldots(12)$$

Substituting

$$\frac{1}{T} + aK = 2\zeta\omega_n$$

$$\text{and } \frac{K}{T} = \omega_n^2 \quad \ldots\ldots(13)$$

Equation (12) becomes ;

$$\ddot{\phi} + 2\zeta\omega_n\dot{\phi} + \omega_n^2\phi = \omega_n^2 \, \_\_\_\_ \quad \ldots\ldots(14)$$

System transfer function will be ;

$$G(s) = \frac{K}{T}\,\frac{1 + aTs}{s^2 + (1/T + aK)s + \frac{K}{T}}$$

Substituting equation (13)

$$G(s) = \frac{\omega_n^2 + \left(2\zeta\omega_n - \frac{\omega_n^2}{K}\right)s}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad \ldots\ldots(15)$$

(i) System frequency response will be ;

$$\frac{\omega_2}{\omega_1} = \frac{1 + j\frac{\omega}{\omega_n}\left(2\zeta - \frac{\omega_n}{K}\right)}{1 + 2j\zeta\frac{\omega}{\omega_n} - \frac{\omega^2}{\omega_n^2}} \quad \ldots\ldots(16)$$

From Equation - 25, it is clear that two extreme values of

$$\frac{\omega_n}{K} \text{ are } \frac{\omega_n}{K} = 2\zeta \quad \text{and } \omega_n/K = 0 \qquad \ldots\ldots(26)$$

for these two values of ω_n/K and $\zeta = .5$, the frequency response is shown in Figure 2.6.

Transient Response :-

System transient response to step input will be ;

$$\phi_2(t) = U(t)\left[1 - e^{-\zeta\omega_n t}\cos\sqrt{1-\zeta^2}\,\omega_n t - \frac{\zeta - \omega_n/K}{\sqrt{1-\zeta^2}}\sin\sqrt{1-\zeta^2}\,\omega_n t\right] \qquad \ldots\ldots(27)$$

Transient response for $\zeta = .5$ and two limiting values ..(26) is shown in Figure 2.5 .

Pull-in-Range :-

Gruen(2) has shown that as $\omega_n/K \rightarrow 0$ the capture range of the system is approximately

$$\Delta\omega \text{ pull in} = \sqrt{2\zeta\,\omega_n K} \qquad \ldots\ldots(28)$$

for $\zeta = .5$

Pull in range to lock range ratio is

$$= \frac{\sqrt{\omega_n K}}{K} = \sqrt{\frac{\omega_n}{K}}$$

For simple R.C. filter

capture range is $= \sqrt{w_n^2}$

Substituting $2 w_n = w_n^2/K$ from (13) ;

$= w_n$

Pull in ratio is $= w_n/K$

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APPENDIX - III

a) System response to F.M. Signal :

When the system input is

$$A_1 \sin (w_1 + \delta \sin w_m t)\, t \qquad \text{....(19)}$$

for $\delta/w_m^2 \ll 1$ and if the carrier frequency of input is the same as the centre frequency of oscillator.

The detuning is function of time

$$\Omega = \delta \sin w_m t \qquad \text{....(20)}$$

Now the steady state solution of the basic differential equation as shown in appendix - I, equation (6.a) is

$$\sin \phi = \frac{\Omega}{K F(jw)}$$

where F(jw) is the filter gain.

Now substituting (2) in it:

$$\sin \phi = \frac{\delta \sin w_m t}{K/F(jw_m)} \qquad \text{.....(21)}$$

Equation (21) shows that phase detector output is a sinusoidal voltage of frequency w_m and amplitude

$$\frac{K_o \delta}{K/F(jw_m)}.$$

It also shows that ϕ is not constant but a function of time. Hence the oscillator output will be phase modulated and modulation will be given by

$$\phi(t) = \sin^{-1}\left(\frac{\delta}{K/V\ (j\omega m)} \sin \omega_m t\right) \quad \text{......(21)}$$

Max. phase modulation will be

$$-\pi/2 \quad \text{to} \quad \pi/2 \quad \text{for filter less case and } \delta = K.$$

Response to A.M. Signal :-

Let the input signal be

$$A_1 (1 + m \cos \omega_m t) \sin \omega_1 t \quad \text{...(22)}$$

As shown in Appendix - I, output of phase detector is $K_o \sin \phi$ where K_o is a function of input voltage. For the switching voltage i.e. oscillator voltage is high compared to input signal phase detector gain K_o is directly proportional to input voltage.

$$K_o = K A_1 (1 + m \cos \omega_m t) \quad \text{.........(23)}$$

As shown in Appendix - I that phase detector output

$$K_o \sin \phi = \Omega / K_2$$

Substituting (22) in Equation (23)

$$K A_1 (1 + m \cos \omega_m t) \sin \phi = \Omega / K^2$$

$$\text{or } (1 + m \cos \omega mt) \sin \phi = \frac{\Omega}{K} \quad \text{.......(24)}$$

$$\text{or } \phi = \sin^{-1} \frac{\Omega}{K(1 + m \cos \omega_m t)} \quad \text{.......(25)}$$

Equation - (25) shows that phase error in the output is function of time.

From equation (24) max. modulation for which system will lock is

$$m_{max} \quad 1 - \frac{\Omega}{K} \qquad \ldots\ldots(26)$$

i.e. for minimum value of input amplitude the system locks at maximum phase error $\pi/2$

Maximum phase modulation in output for amplitude modulation of input given by equation (20) will be ;

$$\pi/2 - \sin^{-1} \frac{\Omega}{K\left(2 - \frac{\Omega}{K}\right)}$$

$$= \pi/2 - \sin^{-1} \frac{\Omega}{2K - \Omega} \qquad \ldots\ldots(27)$$

As Ω tends to zero phase modulation increases but as soon $\Omega = 0$ equation - 22 does not hold good.

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CHAPTER - II

2.1 MODIFIED AFC SYSTEM :

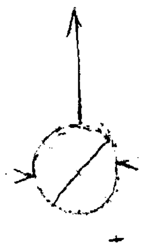
The AFC system as discussed in Chapter - I suffers from frequency error with frequency modulated inputs and total loss of amplitude information etc. For its operation it also inherently requires a finite error. A self adaptive lead lag system which will cause an appreciable reduction in phase error for frequency modulated inputs and which will also reproduce the amplitude information in the original signal will now be discussed in this Chapter. This modified system is indicated schematically in Fig. 2.1.

It is an orthogonal system in which two sets of modulating frequencies 90° apart are employed. It is based on the relation.

$$KA(t)\cos(wt) = K_1K_2\,A(t)\cos(wt+\phi(t))\cos\phi(t) + K_1K_2\,A(t)\sin(wt+\phi(t))\sin\phi(t) \quad \ldots\ldots\ldots\ldots(2.1)$$

where K, K_1, K_2 are constants.

If the input is of the form $A(t)\sin wt$ the ideal V.C.O. output would be of the form $K_1\cos wt$. But there will always be a certain phase error so that the V.C.O. output can be represented by a function of the type $K_1[\cos wt + \phi(t)]$ The error voltage of the phase lock loop is of the form $K_2A(t)\sin\phi(t)$. By employing an additional modulator



is to a phase shifted voltage of the form $K_2 \sin (wt + \phi(t))$ an error function $K_2 A(t) \cos \phi(t)$ can be obtained. This assumes that the two modulators are identical. Then all the terms in equation 2.1 are available. It only remains to perform electronically the multiplication and summation operation indicated in equation 2.1. Here these multiplications $K_1 \cos [wt + \phi(t)]$ to $K_2 A(t) \cos \phi(t)$ and $K_2 \sin (wt + \phi(t))$ to $K_2 A(t) \sin \phi(t)$ are performed by two modulators followed by appropriate tuned filters. Summation is done in an adder to get desired function. The whole operation performed is a self adaptive control which is discussed in the following section.

2.2 SELF ADAPTIVE CONTROL SYSTEM :

The system discussed in the present thesis is one method which employs self adaptive control system i.e. which is inherently capable of maintaining desired performance under changing environment conditions which is changing frequency in this case.

A self adaptive control system should have three essential features :

(a) Identification :- It is the process by which the system is characterised. A system can be defined by its transfer function or impulse response, but identification implies in addition to this, the formation of an index of performance or figure of merit. It is the criteria for controlling the system parameters to yield the desired performance.

In this case equation 2.1 provides the design criteria and decides the adaptivity of the system.

(b) Comparison : In order to correct the system parameters, an adaptive control system should have some means of generating an error voltage by comparing the input and output. This error voltage which correct the system parameter may not be the same as the one that operates the main control loop. In the present case the comparison is made in two orthogonal channels using two phase detectors and error functions sin ϕ and cos ϕ are generated for providing correction according to equation 2.1 .

(c) Adjustment : Finally the system parameters are adjusted by error functions according to index of performance. Adjusted parameters may be gain or time constants. In this case, the correction is applied to gain and phase.

2.3 THEORY OF PHASE LOCK LOOP :

The basic AFC system which is shown in Fig. 1.1 can be represented by its transfer function [8],

$$G(s) = \frac{\omega_n^2}{s^2 + 2\,\omega_n + \omega_n^2}$$

where $\omega_n^2 = (K/T)\cos\phi$ natural frequency of the system

$2\xi\omega_n = 1/T$

ξ = damping coefficient.

Here both φ and v_a are the functions of gain K and phase error ϕ. With the change in frequency i.e oscillator drift or change in input frequency the phase error ϕ is changing hence the system parameters φ and v_a are changing. In the present system which can also be represented as in Figure 2.2, the correction is applied in gain K and phase ϕ so to get the desired output. The desired output of the system is input itself. In Figure 2.2, the computer compares the desired output (which is input itself) to the actual output of AFC system and thereby generates the error functions according to equation 2.1. With the aid of these functions, the correction is applied to the output. It may be noted that this correction is done by an open loop system. This system can be classified as signal synthesis adaptive control system (in which error functions are generated and the correction is applied to the output by open loop control so as to bring it to the desired value. The main feed back loop of the system is not at all affected all the properties of main loop are preserved. So in the present case also, the characteristics and performance of main AFC loop are maintained.

(a) Response to Constant Frequency Input :

In Figure 2.1 if input is A sin wt i.e. a constant frequency input the output of the adder A will be given by relation ;

$$K A \cos(\omega t + \phi) \cos\phi + KA \sin(\omega t + \phi) \sin\phi$$

$$= K A \cos \omega t \qquad \ldots\ldots\ldots\ldots(2.3)$$

where ϕ = the phase error

K = a constant.

Equation 2.3 shows that the system output is not a function of ϕ which is equal to arc sin $\frac{\Omega}{K}$ where Ω is the detuning, so for all the frequencies the phase difference remains 90 degree. Performance of the system that is lock range, capture range etc., are maintained constant as shown in Table No. 2.1, which gives a comparison between simple phase lock system and modification system. It is assumed that the main loop filter characteristics are not affected by filter F_1 and F_2, shown in Figure 2.1 . Equation 2.3 also indicates that the modified system behaves like an exactly tuned linear amplifier while simple APC system behaves like a tuned limiter.

Table 2.1

Comparison of simple APC system to modified system.

	Simple APC System: Without filter	Simple APC System: With filter	Modified system
Lock range	2 K	2 K	2 K
Capture range	2 K	$\sqrt{2\zeta\omega_n K}$	$\sqrt{2\zeta\omega_n K}$
Steady state error			
i) For const. freq. input	$\phi = \sin\frac{\Omega}{K}$	$\phi = \sin\frac{\Omega}{K}$	$\phi = 0$
ii) For f.m. input	$\phi(t) = \sin^{-1}\Omega/K_2 a(t)$	$\phi(t) = \sin^{-1}\Omega/K_2 a(t)$	$\phi(t) = 0$
iii) For A.M. input	$\phi(t) = \sin^{-1}$	$\phi(t) = \sin^{-1}$	$\phi(t) = 0$

Equation 2.3 is derived for steady state condition but in transient state also where the V.C.O. output is frequency modulated and the loop error is a varying voltage, system error is neutralised.

(b) Response to f.m. Input :

In Figure 2.1 if the input is frequency modulated that is A cos $(W(t)\ t)$ where frequency is a function of time, from table 2.1, the loop error $\phi(t)$ is equal to Arc sin $\frac{w(t) - w_0}{K.F}$ where w_0 is the centre frequency of the oscillator and F is the filter gain. For ramp modulated inputs that is for input A cos $(w + Bt)t$ the error $\phi(t)$ goes on increasing with time till the system unlocks. In Figure 2.1, the output of the multiplier M_1 and M_2 are

$$e_{m1} = K\ A \sin \{ w(t)\ t + \phi(t) \} \sin \phi(t) \quad .....(2.4)$$

$$e_{m2} = K.A \cos \{ w(t)t + \phi(t) \} \cos \phi(t) \quad ......(2.5)$$

and the output of the adder which is summation of equations 2.4 and 2.5 will be

$$e_o = K\ A \cos \{ w(t)\ t \} \quad ....(2.6)$$

In deriving equation (2.6) assumptions made are, one, that the two channels are perfectly orthogonal, second, the gains of the two channels are equal. So in this case with these two assumptions the output of the system as shown by equation 2.6 is the same as the input with no phase or amplitude error. The adaptive system used for modifying the control mean system being an open loop system.

it does not affect the operation of the phase lock loop and the locking conditions remains the same as explained in Sec. 2.3 that $\delta < \omega_n^2/K$ [8] where δ is the modulation index, ω_n is the natural frequency and K is the system open loop gain.

(c) Response to A.M. Input :

From Table 2.1 the phase error in the output $\phi(t)$ which is function of time is equal to Arc sin $\frac{\Omega}{K_2 A(t)}$ where A(t) is the modulated input amplitude and K_2 is constant. The correction voltage from the orthogonal channel will then be A(t) - Sin $\phi(t)$, which is a varying voltage with time. Then the system out put will be the summation of two multipliers output where correction is applied. The corrected out put will be

$$e_o = K\,A\,(t) \quad \text{Cos}\ \omega t \quad ....(2.9)$$

Equation 2.9 is valid with the assumptions that the two channels are perfectly orthogonal, there gains are equal and the filters are having flat characteristics for the frequency variation of correction voltage. The output given by equation 2.9 consists of all amplitude informations present in the input.

Permissible depth of modulation for which the oscillation will remain locked with the input depends upon initial detuning of the carrier frequency and is not affected by this open loop modification. Even at the bottom of the modulation if the oscillator tends to unlock no serious error is involved, because equation (2.0) will hold true, so long as $\phi(t)$ has a frequency component with in the band width of low pass filter.

Sec. 2.4 <u>Imperfect System</u> :

It is assumed in the preceding section that

1. The two channels are perfectly orthogonal that is signals in the two channels are correctly at 90° phase difference.
2. The gains of the two channels are exactly equal.

But in practice both phase and gain are subjected to experimental error. With these errors the system behaviour will not be exactly the same as discussed in preceding section. There will be some amplitude and phase error in the output depending upon the difference of the two gains, departure from orthogonality and the initial detuning of the oscillator.

(a) Response for Constant Frequency Input:

1. When the signals in two channels are at a phase difference $90 \pm \epsilon$ instead of 90°. Where ϵ is so small as $\sin\epsilon$ can be approximated to ϵ and $\cos\epsilon$ to $1 - \epsilon^2/2$. If ϵ is .1 radians then $\cos\epsilon$ will be equal to .995 which is nearly equal to unity. So for 6 degree to 7 degree phase error in two channels, the output of the

system is almost the same as the input with no phase or amplitude error (vide appendix III).

ii) When the two channels are orthogonal, but there is a small difference in two gains such as gain of one channel is K while the gain of the other channel is K (1 - x). The system output will have some phase and amplitude error. These errors

$$\theta = \frac{x}{2} \sin 2\phi \qquad ......(2.10)$$

$$y = x \sin 2\phi \qquad ......(2.11)$$

are plotted in Figure 2.3 and 2.4, for different values of x, y are the phase and amplitude error in system output. θ and the phase error in the oscillator output which is equal to Arc Sin $\frac{\Omega}{K}$ where Ω is the initial detuning. Table 2.2 contains phase and gain error in the output. For x equal to .1 or less these errors are not appreciable but as x increases, these errors also increase correspondingly.

TABLE - 2.2

Phase and Amplitude error for different gain errors.

	Max. phase error	Max. error $\Delta_1 = \Delta_0/A$
x = .05	± 1.5 degree	.05
x = .1	± 3 degree	.1
x = .15	± 4.5 degree	.15
x = .2	± 6 degree	.20
x = .25	± 7.5 degree	.25
x = .3	± 9 degree	.30

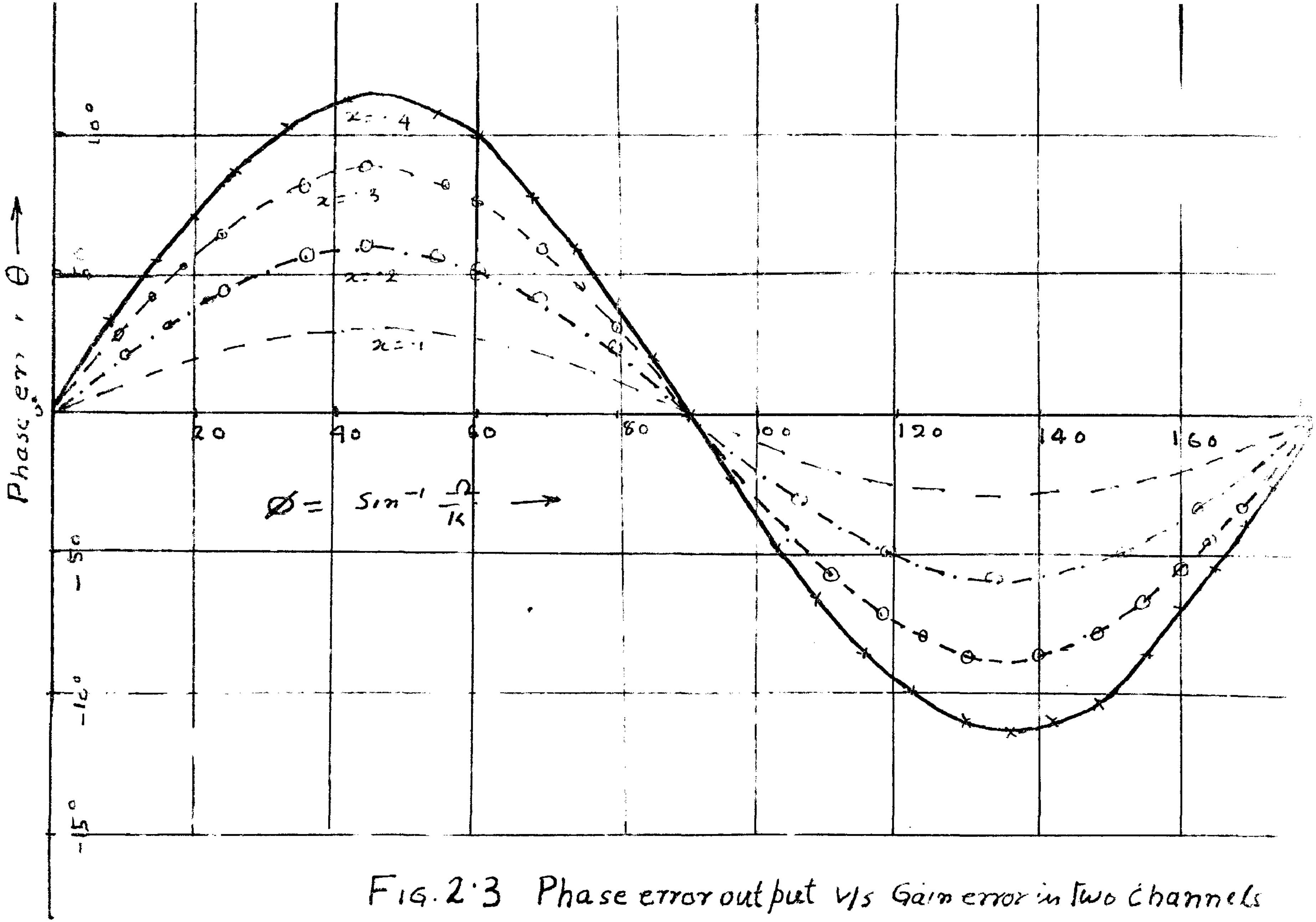
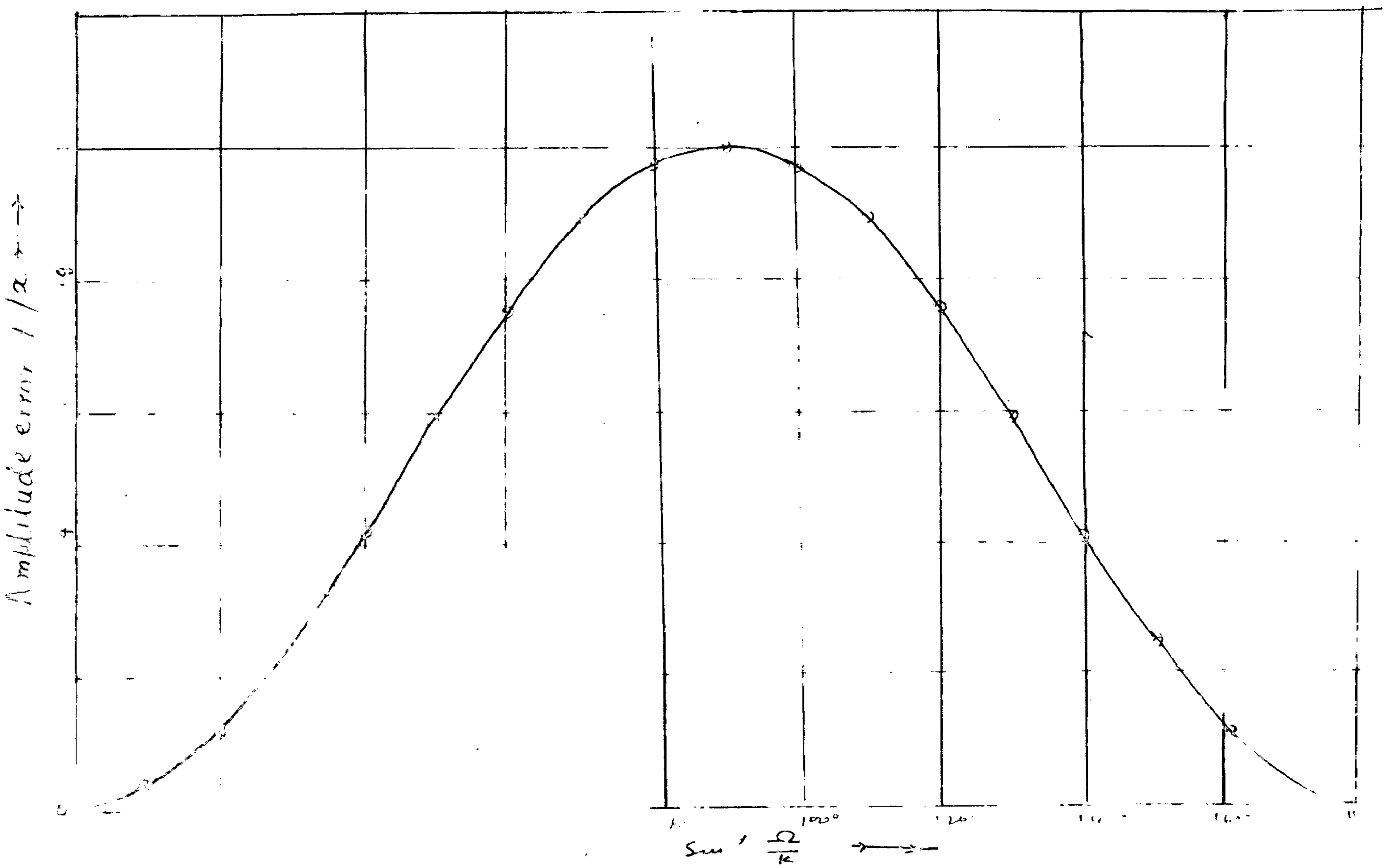


FIG. 2·3 Phase error output v/s Gain error in Two Channels



lised Amplitude

D. RESPONSE TO F.M. ERRORS:

From the similar treatment as done for constant frequency input it can be shown that small phase shifting error is not much affecting the out put , but for small gain error, the output will have some phase and amplitude modulation. Maximum modulations for different gain errors are shown in Table 2.3 (vide Appendix III).

TABLE 2.3

Output phase and amplitude modulation for different gain errors.

δ/π	m =	Max. phase modl.	Max.amp. modl.
	m = .1	±2.7°	.023
.5	m = .15	± 4.1°	.037
	m = .20	± 5.4°	.05
	m = .25	± 6.8°	.0633
	m = .30	± 8.1°	.075

As indicated by table 2.3, for small amplitude errors the phase and amplitude modulation are quite small and can be tolerated in almost all practical purposes.

RESPONSE TO A.M. INPUT :

For small phase shifting error the system output will follow the input with negligible errors. But for gain error in the two channels, the output will have some phase modulation and extra amplitude modulation. The output amplitude for $\Delta(t) \sin \omega t$ input will be

$$|e_{om}| \simeq n\, A(t) \left[1 \mp n \sin 2\phi \right] \quad \ldots\ldots(2.15)$$

where $$\phi(t) = \sin^{-1} \frac{\Omega}{n \frac{\Delta(t)}{A}} \quad \ldots\ldots\ldots(2.16)$$

(vide Appendix III).

where A is carrier amplitude

Ω is frequency difference between carrier and oscillator centre frequency.

Equation (2.16) shows that output is having an amplitude distortion , and this distortion is proportional to amplitude modulation and the gain error. For $n = .1$ maximum distortion possible for 100% of modulation will be 10% which is appreciable.

The phase of the output is given by

$$\Theta = \tan^{-1} \frac{n}{2} \sin 2\phi(t) \quad \ldots\ldots(2.17)$$

Equation 2.17 shows that the output will also be phase modulated also and for $n = .1$ max. phase modulation will be .05 and which is negligible.

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APPENDIX - I.

(a) Modified APC system response to constant frequency input :

For constant frequency input i.e.

$$e_i = A_i \sin w t \qquad \ldots\ldots\ldots\ldots 1$$

when the system is locked the out puts of phase detectors are d.c. voltages

$$ed_1 = K_o A_i \sin \phi \qquad \ldots\ldots\ldots\ldots (2a)$$

$$ed_2 = K_o A_i \cos \phi \qquad \ldots\ldots\ldots\ldots (2b)$$

and the phase locked oscillator output will be

$$e_o = A_o \cos (wt + \phi) \qquad \ldots\ldots\ldots\ldots (3)$$

Now the out put of multipliers will be :

$$em_1 = K_o A_i A_o \cos (w t + \phi) \sin \phi \qquad \ldots (4a)$$

$$em_2 = K_o A_i A_o \sin (w t + \phi) \cos \phi \qquad \ldots\ldots (4b)$$

Hence the system output that sum of 4-a and 4 b will be

$$e_s = K_o A_i A_o \cos w t \qquad \ldots\ldots\ldots\ldots (5)$$

(b) System response when oscillator is not locked.

When the oscillator is not locked the outputs of the phase detectors will be a.c. voltage of frequency equal to the difference of oscillator frequency and input frequency

$$e d_1 = K_o A_i \sin (w_o - w_i) t \qquad \ldots\ldots\ldots (6-a)$$

$$e d_2 = K_o A_i \cos (w_o - w_i) t \qquad \ldots\ldots (6-b)$$

where w_o = oscillator centre frequency

and w_i = input signal frequency

putting $w_o - w_i = \Omega$ ......(7)

e(a) and e(b) are

$$e_1 = E_o A_i \sin \Omega t \qquad .... (8-a)$$

$$e_2 = E_o A_i \cos \Omega t \qquad .... (8-b)$$

Now the output of the multipliers will be

$$e_{m1} = \frac{E_o A_i}{F(j\Omega)} \cos \Omega t \cos w_o t \qquad ..... (9-a)$$

$$e_{m2} = \frac{E_o A_i}{F(j\Omega)} \sin \Omega t \sin w_o t \qquad ..... (9-b)$$

where $F(j\Omega)$ is the filter transfer function.

Now the output of the adder will be :

$$e_A = \frac{E_o A_i}{F(j\Omega)} \cos \left[(w_o - \Omega) t + \angle F(j\Omega)\right]$$

.......(10)

From (7) and (10)

$$e_A = \frac{E_o A_i}{F(j\Omega)} \left[\cos w_i t + \angle F(j\Omega)\right] \qquad .... (11)$$

-oOo-

APPENDIX - II.

(a) Modified AFC System Response to F.M. Input :-

In 2.1 if the input is

$$e_i = A_i \sin\left[w_i + \quad \sin w_m t\right] t \qquad \text{......(12)}$$

where A_i is the carrier amplitude and   is the modulation index assuming carrier frequency to be the same as oscillator centre frequency, the phase locked oscillator output will be ;

$$e_o = A_o \cos\left[\left(w_i + \quad \sin w_m t\right) t + \phi(t)\right] \qquad \text{......(13)}$$

where $\phi(t) = \sin^{-1} \dfrac{\delta}{K\, F(jwm)}$

Output of the phase detector D_1, D_2 will be

$$ed_1 = K_o \sin \phi(t) \qquad \text{......(14-a)}$$

$$ed_2 = K_o \cos \phi(t) \qquad \text{......(14-b)}$$

Multiplier outputs will be

$$em_1 = \frac{K_o A_o}{F(jwm)} \sin \phi(t) \sin\left(w_i + \sin w_m t\right) t + \phi(t) \qquad \text{......(15-a)}$$

$$em_2 = \frac{K_o A_o}{F(jwm)} \cos \phi(t) \cos\left(w_i + \sin w_m t\right) t + \phi(t) \qquad \text{......(15-b)}$$

Output of the adder will be ;

$$e = e_{m1} + e_{m2} \qquad \text{......(16)}$$

$$e = \frac{K_d A_1}{F(jw_m)} \cos (w_1 t + \beta \sin w_m t) \qquad ......(17)$$

If the filter are having flat characteristics for the frequency range in which f_m varies, the output is having same frequency modulation as the input signal and the amplitude is linearly proportional to input.

(b) Response to A.M. Input :-

If the system, input is ;

$$e_1 = A_1 (1 + m \cos w_m t) \sin w t \qquad ......(18)$$

Output of phase lock system will be ;

$$e_o = A_o \cos (wt + \phi (t)) \qquad ....(19)$$

$$\text{where} \quad \phi(t) = \sin^{-1} \frac{\Omega}{K(1 + m \cos w_m t)} \qquad .....(20)$$

where m = modulation index

K = the system gain

w_m = modulation frequency.

Output of the two detectors will be ;

$$e_{d1} = K_d (1 + m \cos w_m t) \sin \phi (t) \qquad .....(21\text{-a})$$

$$e_{d2} = K_d (1 + m \cos w_m t) \cos \phi (t) \qquad .....(21\text{-b})$$

Now the two multipliers output will be ;

$$e_{m1} = K_d A_o (1 + m \cos w_m t) \sin \phi (t) \cos \phi (t) \qquad ....(22a)$$

$$e_{m2} = K_d A_o (1 + m \cos w_m t) \cos \phi(t) \cos (wt + \phi(t)) \qquad ....(22\text{-b})$$

Hence the output of the adder i.e., the system output will be

$$e = e_{m1} + e_{m2} \qquad \text{......(23)}$$

$$e = K_o A_o (1 + m \cos w_m t) \cos w t \qquad \text{...(24)}$$

Equation 24 shows that depth of modulation in output is the same as it was in the input and the ouput is not having any phase error.

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APPENDIX - III

If the two channels are not perfectly orthogonal i.e. the signals in two channels are at $(90 \pm \epsilon)$ phase difference instead of 90°, or if the gains of the two channels are not equal, the system will have some phase and amplitude error.

(a) Response to Constant Frequency Input :

If input is

$$e_i = A_i \sin wt \qquad \ldots.(25)$$

output of the phase locked oscillator will be

$$e_o = A_o \cos (wt + \phi) \qquad \ldots\ldots\ldots(26)$$

Now if phase shifting circuit output is

$$e_p = A_o \sin (wt + \phi \pm \epsilon) \qquad \ldots\ldots\ldots\ldots(27)$$

then the two phase detectors out puts will be

$$ed_1 = K_o A_i \sin \phi \qquad \ldots\ldots\ldots\ldots\ldots(28\text{-}a)$$

$$ed_2 = K_o A_i \cos (\phi \pm \epsilon) \qquad \ldots\ldots\ldots\ldots(28\text{-}b)$$

The output of the multipliers

$$e_{m1} = K_o A_i \sin \phi \sin (wt + \phi \pm \epsilon)$$

$$= K_o A_i \sin \phi \{\sin(wt + \phi) \cos \epsilon \pm \sin \epsilon \sin (wt + \phi)\} \qquad \ldots\ldots\ldots(29)$$

Now approximating for small θ,

$$\sin \theta \simeq \theta$$

$$\text{and } \cos \theta = 1 - \frac{\theta^2}{2}$$

for θ to be .1 radian.

$$\cos = 1 - \frac{.01}{2} = .995$$

which very closely unity hence for this error with in ± .1 radian cos can be approximated equal to unity.

So now putting it in equation (30),

$$q_{m1} = K_o A_1 \sin \phi \left\{ \sin (wt + \phi) \pm \theta \cos (wt + \phi) \right\} \quad \text{.......(30-a)}$$

$$\text{and } q_{m2} = K_o A_1 \cos (wt + \phi) \left\{ \cos \phi \mp \theta \sin \phi \right\} \quad \text{..........(30-b)}$$

Adding equations (30-a) and (30-b),

output of the adder will be,

$$e = K_o A_1 \cos (wt + \phi) \quad \text{....................(31)}$$

ii) If the gains of the two channels are K and K (1 ± m)

Output of the multipliers will be,

$$q_{m1} = K \cos (wt + \phi) \cos \phi \quad \text{........... (32-a)}$$

$$q_{m2} = K (1 \pm m) \sin (wt + \phi) \sin \phi \quad \text{....(32-b)}$$

Now adding equations (32-a) and (32-b), the output of the adder,

$$e_\Delta = E\cos wt \pm m\sin(wt+\phi)\sin\phi$$

$$= E\cos wt \pm \frac{Em}{2}\left\{\cos wt - \cos(wt+2\phi)\right\}$$

$$= E(1 \pm m/2)\cos wt \mp \frac{Em}{2}\cos(wt+2\phi)$$

...........(33)

From equation (33) output amplitude will be :

$$e_\Delta = E\sqrt{(1 \pm m/2)^2 + (m/2)^2 \mp m(1 \pm m/2)\cos 2\phi}$$

.......(34)

Neglecting the square terms

$$|e_\Delta| = E\sqrt{1 \pm m \mp m\cos 2\phi} \qquad ....(35)$$

$$E(1 \pm m\sin 2\phi) \qquad ....(36)$$

Phase of the output will be

$$\angle e_\Delta = \tan^{-1}\frac{\pm\, {}^{m}/_{2}\,\sin 2\phi}{(1 \pm {}^{m}/_{2}) \mp {}^{m}/_{2}\,\cos 2\phi}$$

$$\pm (m/2)\sin 2\phi \qquad ....(37)$$

(b) Response to F.M. Input :

(i) For small error in orthogonality of two channel the system output can be approximated to be the same as input with some gain as it is done for constant frequency inputs.

(ii) For the two gain ratios $(1 \pm m)$.

If input is $e_1 = A_1\sin\left\{(w_1 + \sin w_m t)\,t\right\}$ ......(38)

Output of the adder will be :

$$e_A = K \left\{ (1 \pm m) \cos \left\{ (\omega t + \sin \omega_m t)\, t \right\} \right.$$

$$= K\,{}^{m}/_{2} \left\{ (\omega + \sin \omega_m t)\, t + 2\, \phi(t) \right\} \quad .....(59)$$

where $\phi(t) = \sin^{-1} \dfrac{\delta}{K\ F\,(j\omega)}\ \sin \omega_m t$ ....(60)

From equation (59), amplitude of the output will be:

$$e_A = K\,(1 \pm m \sin^2 \phi\,(t)) \quad ....(61)$$

Putting (60)

$$|e_A| = K \left\{ 1 \pm m \left(\frac{\delta}{K\ F(j\omega)} \sin \omega_m t \right)^2 \right\}$$

.........(62)

Equation (62) shows that the output is amplitude modulated.

Phase error in the output will be

$$\angle e_A = \tan^{-1} \frac{{}^{m}/_{2}\ \sin 2\phi}{(1 + m/2) - {}^{m}/_{2} \cos 2\phi}$$

$$\simeq {}^{m}/_{2}\ \sin 2\phi\,(t)$$

$$\simeq m \sin \phi\,(t) \quad ....(63)$$

$$\simeq m \frac{\delta}{K\ F\,(j\omega)}\ \sin \omega_m t \quad ....(64)$$

Example:

For $\delta/K = 1/2$

and $|F\,(j\omega)| = 1$

If gain error m = .05

Max. amplitude modulation in the output will be from equation

$$y(t) = .0125$$

or $= 1.25\ \%$

Max. phase modulation will be from equation

$$\theta(t) = .05 \div 2 = .025 \text{ radian}$$
$$= 1.25^{\circ}$$

For m = .1

Max. amplitude modulation is

$y(t) = .1/4 = .025 = 2.5\ \%$

Max. phase modulation is

$\theta(t) = 1/2 = .05$ radian $= 2.7^{\circ}$

For m = .15

$y(t) = .037$ or $3.7\ \%$

$\theta(t) = 4.1^{\circ}$

(c) Response to A.M. Input ;

(i) For the error in the orthogonality in the channel with in ± .1 radians, the out put will be approximately same as the input.

(ii) For error in two gains i.e. gain of one channel is M while of other is N (M ≥ N). After output will be ;

$$E(t)\cos wt \pm E(t)\,{}^{m}/_{2}\cos(wt + 2\phi(t)) \quad \ldots\ldots\ldots(45)$$

where $$\phi(t) = \sin^{-1}\frac{\Omega}{E(1 + m\cos w_m t)} \quad \ldots\ldots\ldots(46)$$

where $$E(t) = E(1 + m\cos w_m t) \quad \ldots\ldots\ldots\ldots\ldots(47)$$

Now from equation (45) output amplitude is

$$e_0 = E(t)\left\{ 1 \pm m\sin^2\phi(t) \right\} \quad \ldots\ldots\ldots(48)$$

Hence maximum modulation error is

$$= \pm m$$

Phase error in the output will be

$$\theta_1 = \tan\frac{1 \pm {}^{m}/_{2}\sin 2\phi}{1 \pm m\sin^2\phi} \quad \ldots\ldots\ldots(49)$$

$$= \tan^{-1} \pm {}^{m}/_{2}\sin 2\phi \quad \ldots\ldots\ldots(50)$$

Maximum phase modulation in the out put will be

$$\tan^{-1} m/2$$

$$m/2 \quad \ldots\ldots\ldots\ldots\ldots(51)$$

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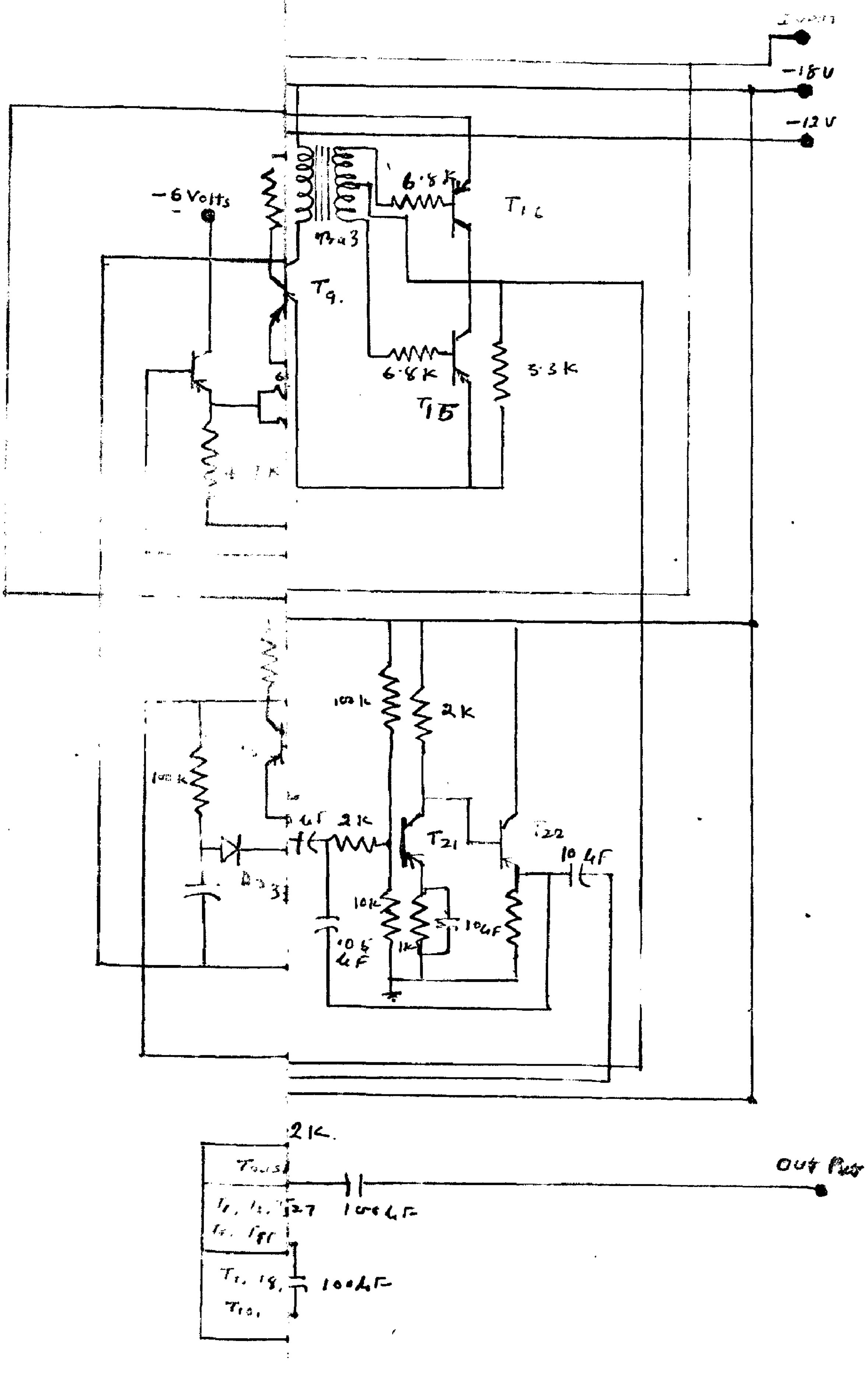
CHAPTER - III

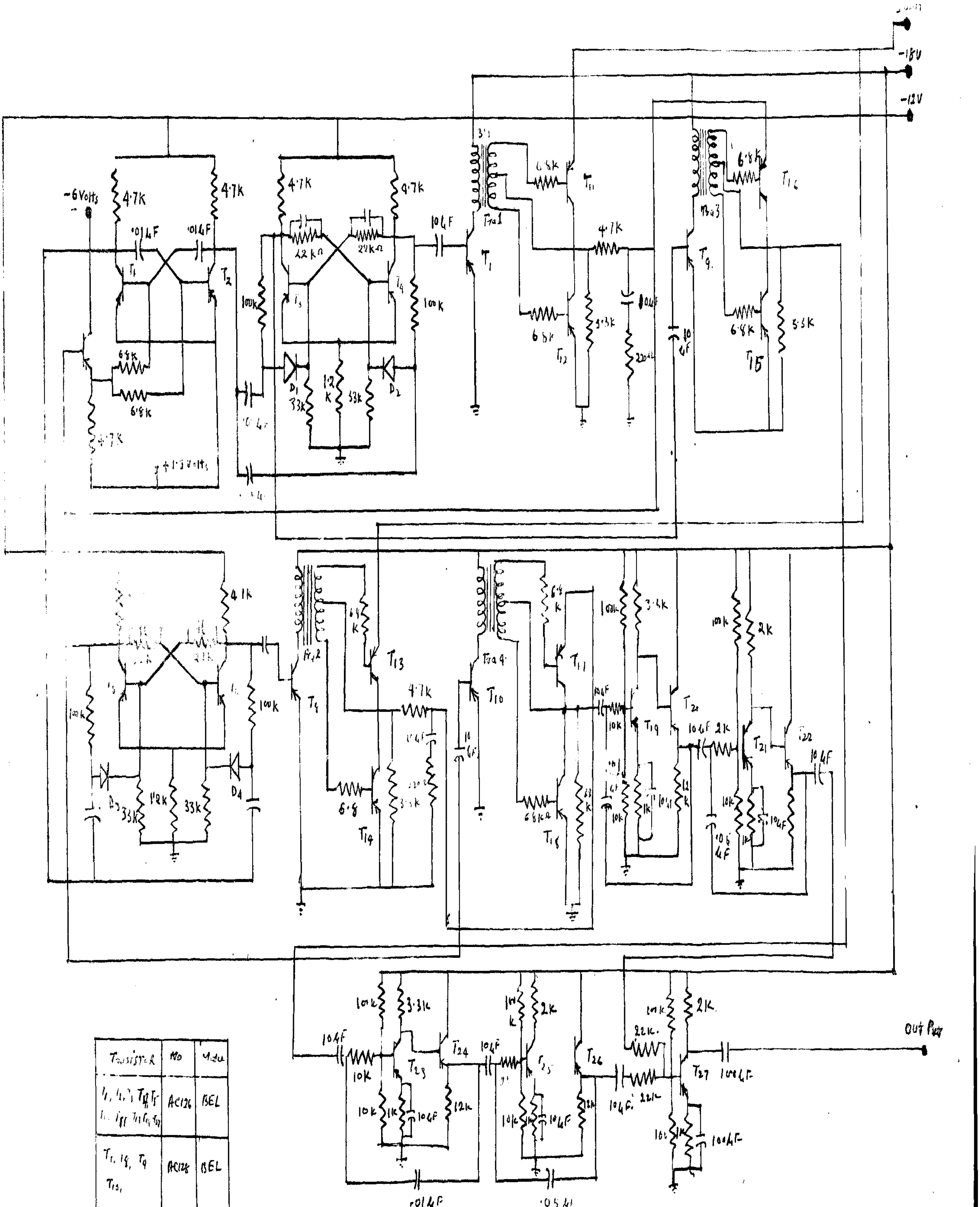
Sec. 3.1 DISCUSSION :

The electronic circuit for the performance of all operations of Figure 2.1 is shown in Figure 3.1 . An astable multi vibrator (Transistors T_1, T_2) functions as voltage controlled oscillator. The frequency is controlled by varying the bias to the multi vibrator by applying the error voltage at point D. In order to obtain stable operation the phase detector is coupled to the astable multivibrator through an emitter follower. The astable multivibrator follower is designed to operate between 1000 c/s to 2000 c/s approximately and its two out puts which are at 180° phase difference are used to trig er two binaries. These (Transistors T_3, T_4, T_5, T_6) binaries yield two outputs 90° apart at a frequency equal to half that of the multi vibrator. That is they vary in frequency between approximately 500 to 1000 cycles as indicated in Figure 3.2 . As the input to the two binaries are 180° apart, their output remains constant at 90° apart over the entire range of operation.

Outputs of the two binaries are used in the two orthogonal channels for switching balanced d.c. chopper circuits (Transistors T_{11}, T_{12}, T_{13} , T_{14}) which act as phase detectors. Buffer stages (Transistor T_7, T_8) are used to avoid the loading of binaries. The various modulation components from the phase detectors



FIED APC SYSTEM



Transistor	No	Make
T1, T2, T3, T4, T5, T6, T11, T12, T13, T14, T17 [illegible]	AC126	BEL
T1, T8, T9, T15,	AC128	BEL

FIG. 3.1 :- CIRCUIT DIAGRAM OF MODIFIED APC SYSTEM

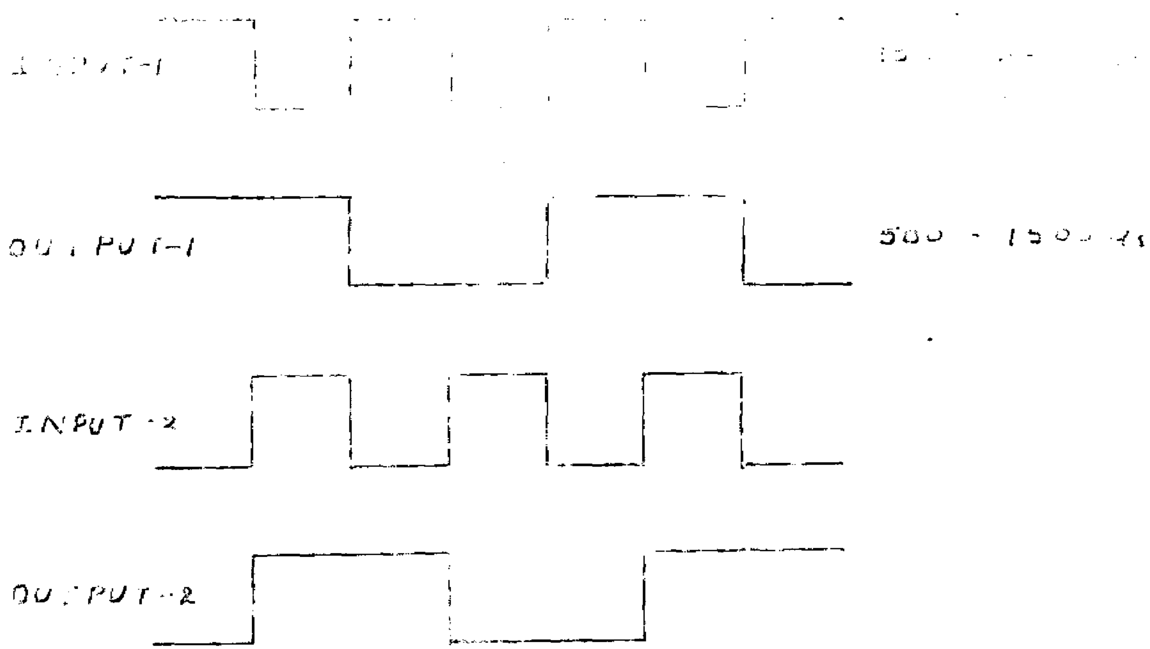


FIG 32. 90° Phase shifting using Binaries

applied to low pass filters so that only the two difference frequency components remains. These low frequency compo- nents (100 c/s) forms the voltages $K_b A_1 \sin \phi(t)$ and $K_b A_1 \cos \phi(t)$. The multiplication of $K_b A_1 \cos \phi(t)$ with $A_o \cos (wt + \phi(t))$ and of $K_b A_1 \sin \phi(t)$ with $A_o \sin (wt + \phi(t))$ is performed in the chopper circuits. (T_{15} T_{16}, T_{17}, T_{18}) respectively. The choppers are identi- cally the same as the ones for obtaining the phase difference signals. The harmonic components are again filtered out from the multiplier out puts using two pairs of active integrators (Transistors T_{19}, T_{20}, T_{21} T_{22}, T_{23}, T_{24}, T_{25}, T_{26}). They are then added up together using an adder circuit (Transistor T_{27}).

Sec. 3.2 MEASUREMENTS :

Measurements are taken using a sinusoidal input through an amplifier. Input voltage was kept fixed such as the amplifier out put was 2 volts. The measurements are taken for phase detector gain, V.C.O. gain, lock range and capture for simple A.P.C. system and modified system. The results are tabulated and shown in Table 3.1 .

V.C.O. centre frequency is 1150 c/s for frequencies between the range, 1030 to 1250 c/s the system locks in and once the system is locked, the frequency could be varied to 1000 cycles after that system unlocks.

TABLE - 3.1

A.P.C. System Performance.

	A.P.C. loop	Modified system
Phase detector gain	.39	-
V.C.gain	$2\pi \times 337$ rad/sec/volt	-
Open loop gain K	$2\pi \times 491.2$ rad/sec	-
ω_n	520 rad/sec	-
ζ	.39	-
Lock range	± 600 c/s	± 600 c/s
Capture range	104 c/s	104 c/s

On the other side if frequency is decreased, it can be varied to 900 c/s and after that system unlocks. This unsymmetrical characteristics is noticed due to non-linear characteristics of the astable multi vibrator. For positive voltage at point D, reaching to 1.5 volts, the multi vibrator characteristics is highly non linear. This effect can be modified by biasing the multi-vibrator circuit to a larger negative voltage but in that case the sensitivity of V.C.O. reduces considerably.

The two outputs of the phase detectors E_oA_2 cos ϕ and E_oA_2 sin ϕ are measured for different frequencies varying from 1000 - 1030 c/s and they are plotted as shown in Figure 3.3. The figure shows that these output vary very nearly as the cosine and sine functions of phase angle error. It is also seen from this plot that the two channels are very closely orthogonal.

Output of the adder was compared with input on C.R.O. screen for different frequencies and the variation of phase difference with frequency was not noticeable. The amplitude frequency response is shown in Figure 3.4. The variation from the critical constant value is due to unbalanced error in the gain of the two channels.

The input v/s output amplitude characteristics is shown in Figure 3.5. It may be noted that the variation is linear to a very high degree of accuracy, but that some slight change occurs when signal frequency is varied.

Due to shortage of time and lack of equipment dynamic response curves with amplitude and frequency modulated signals could not be performed, but the static characteristics given above indicates the great improvement that is obtainable for using the system, as it adds to fidelity in responding to input signals.

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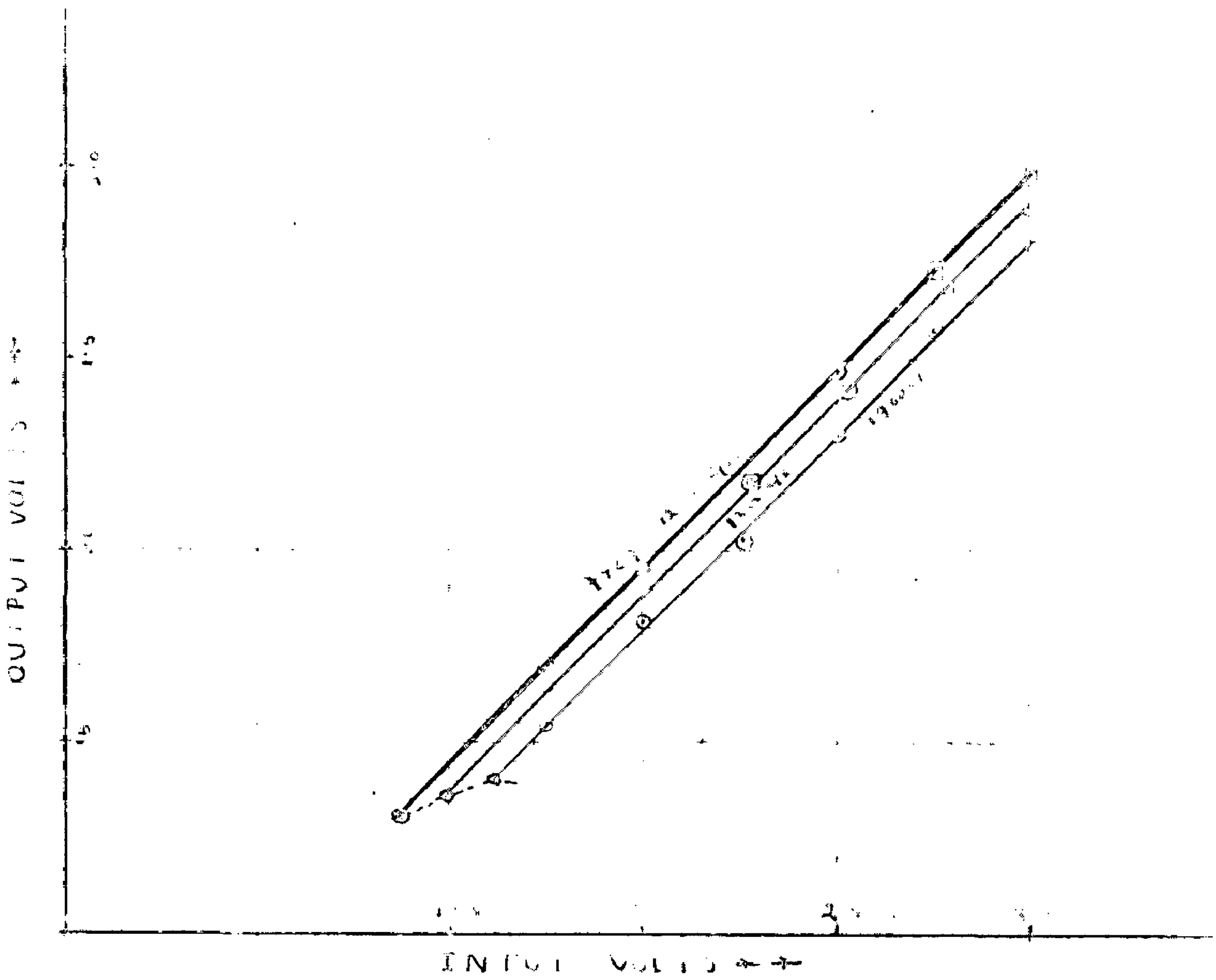


FIG 3.5 – OUTPUT AMPLITUDE VARIATION

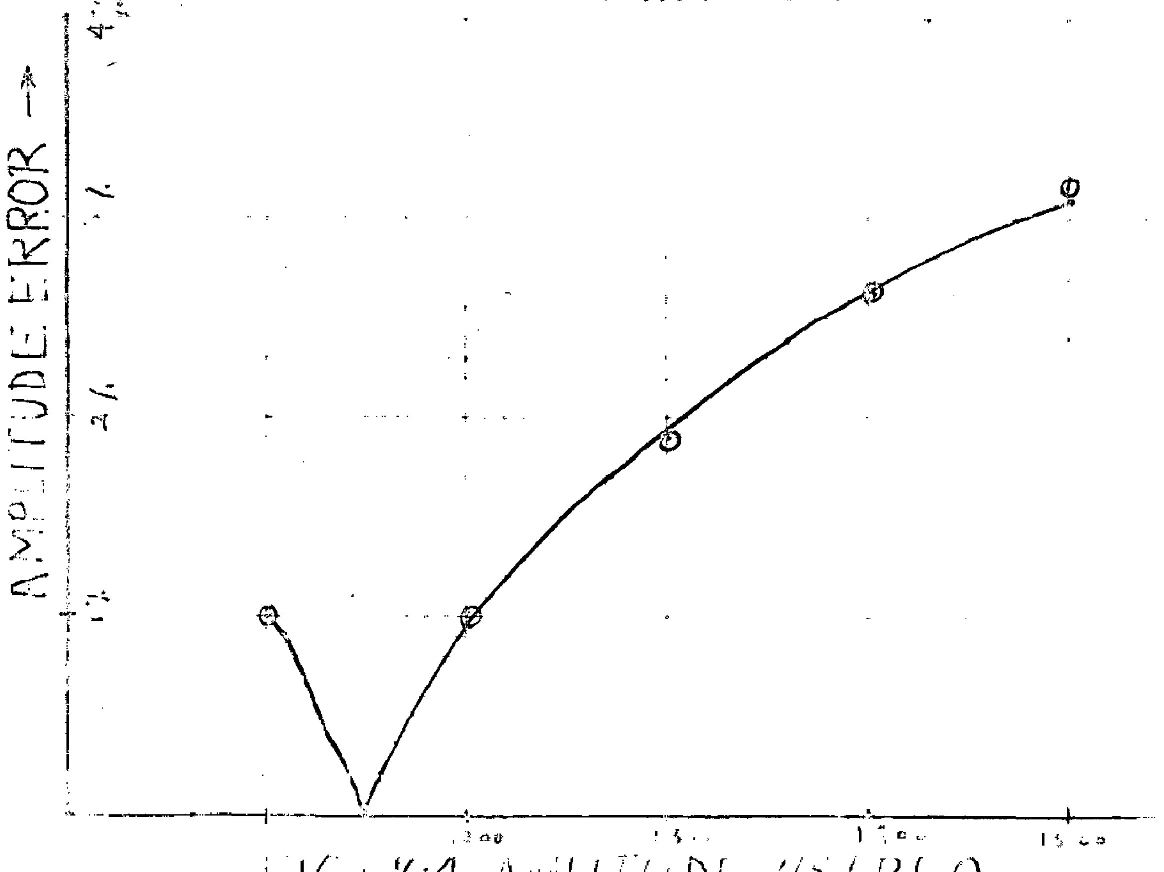
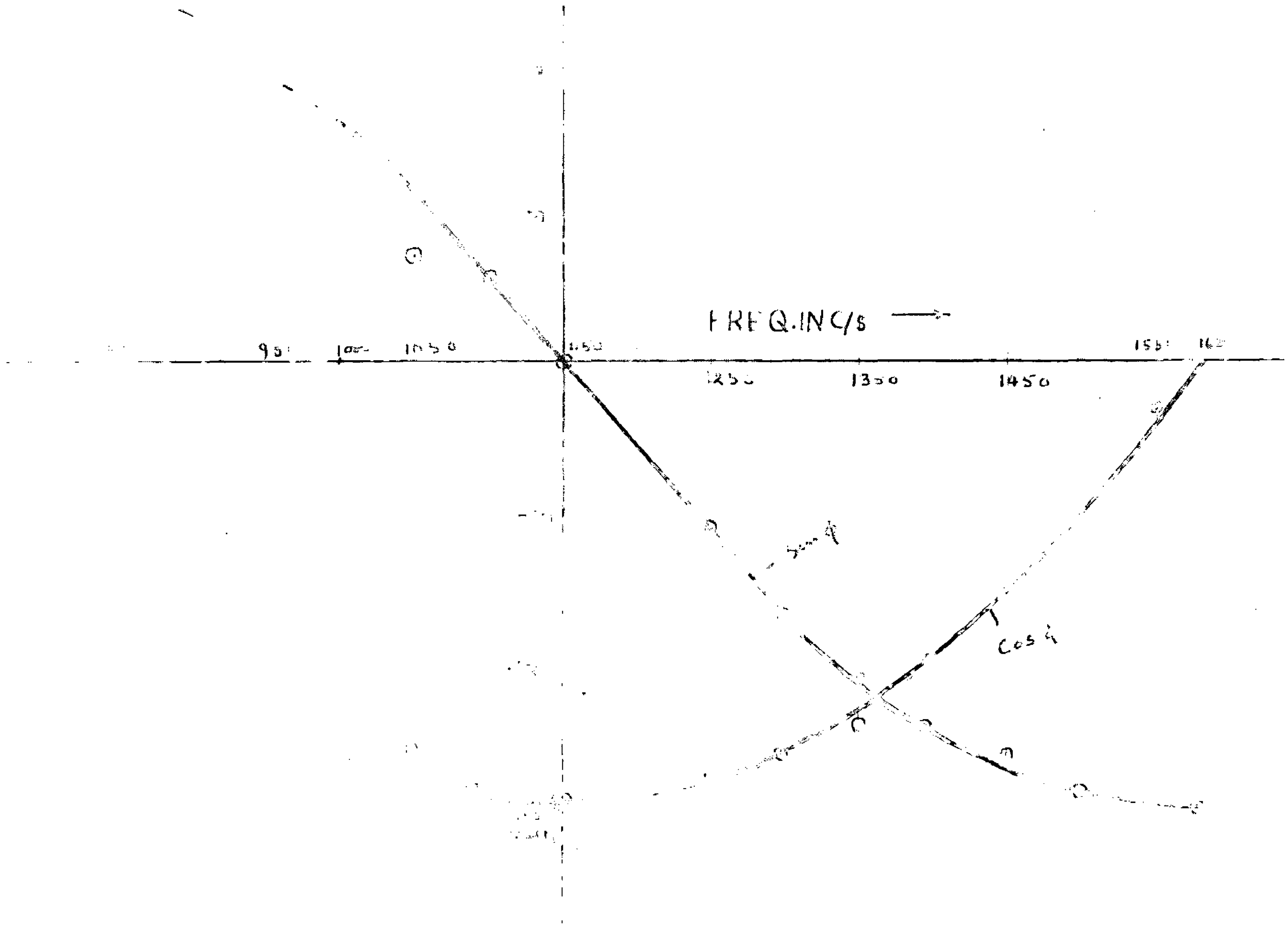


FIG 3.4 AMPLITUDE VS FREQ.


FREQ.IN C/S
95
100
1050
1150
1250
1350
1450
1550
Sin φ
Cos φ

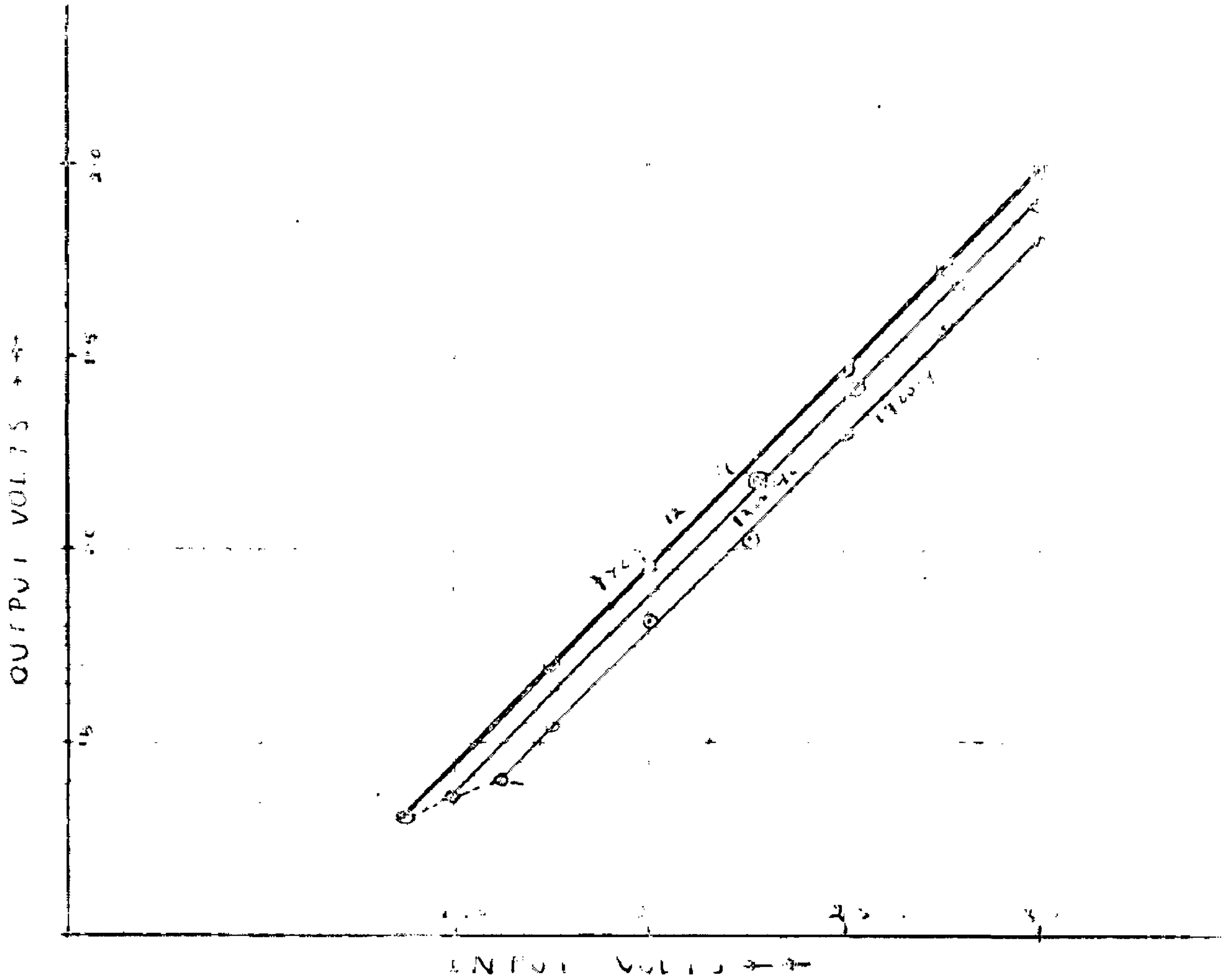


FIG 3.5: OUTPUT AMPLITUDE VARIATION

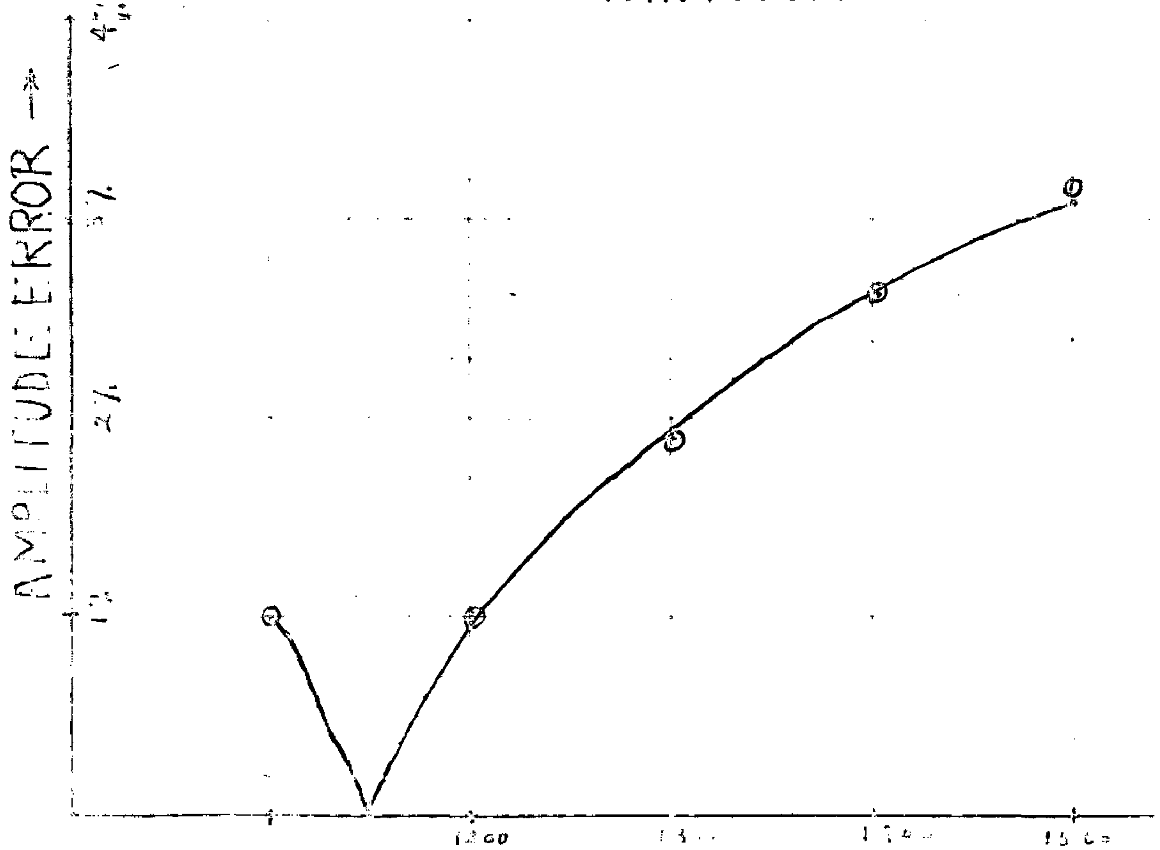


FIG 3.4 AMLITUDE V/S FREQ

<u>Conclusions</u>

The method described here to suppress the Phase error inherent to automatic phase lock systems utilises the simple relation.

$$a \cos wt = a \cos \phi \, . \cos (wt + \phi) + a \sin \phi \sin (wt + \phi) .$$

where ϕ is the phase error of the phase lock system whose output is of the form $\cos (wt + \phi)$,. when the input is of the form $a \cos wt$. The $\sin (wt + \phi)$ is obtained simply by 90° phase rotation of the phase locked oscillator voltage and the $a \cos \phi$, $a \sin \phi$ terms are obtained by the phase detectors. The system which has been experimentally verified, therefore retains both the phase information and the amplitude information of the input signal whereas in the simple phase lock system the amplitude information is totally destroyed and the phase information is subject to certain errors unless the tuning has to be exact. The disadvantage of the system is that the two parallel channels are needed and they have to be balanced separately both in phase and in gain. However the design stringency is considerably reduced by the use of negative feedback. This becomes possible because both the input and output are ideally identical in form. So, in effect, the system is a self tuning narrow band amplifier. Any error introduced in the two channels is of tuning distortion produced at the output and can be suppressed in the usual manner by negative feedback. This was also found true experimentally.

with the resources available only the principle has been verified . A practical system would require the use of phase lock systems of greater precision than that used here. This entails a larger loop gain, which is possible only by introducing an additional intermediate frequency in the phase lock loop. Even with the simple phase lock system employed here the system indicates good promise for tracking Doppler shifted signals from satellites, for increasing the stability of F M systems, pitch and formant tracking in vocoder etc.

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